

Modelling of Crosstalk and Delay for Distributed RLCG On-Chip Interconnects For Ramp Input

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Abstract--In order to accurately model high frequency effects, inductance has been taken into consideration. No longer can interconnects be treated as mere delays or lumped RC networks. In that frequency range, the most accurate simulation model for on-chip VLSI interconnects is the distributed RLC model. Unfortunately, this model has many limitations at much higher of operating frequency used in today's VLSI design. The reduction in cross-sectional dimension leads to more tightly couple interconnects and therefore, a higher probability of unwanted crosstalk interference. This can lead to inaccurate simulations if not modelled properly. At even higher frequency, the aggressor net carries a signal that couples to the victim net through the parasitic capacitances. To determine the effects that this crosstalk will have on circuit operation, the resulting delays and logic levels for the victim nets must be computed. This paper proposes a difference model approach to derive crosstalk and delay in the transform domain. A closed form solution for crosstalk and delay is obtained by incorporating initial conditions using difference model approach for distributed RLCG interconnects. The simulation is performed in 0.18 μ m technology node and an error of less than 1% has been achieved with the proposed model when compared with SPICE.

Keywords--Crosstalk Model, Distributed RLCG Segment, Interconnect, Noise, Delay, VLSI

I. INTRODUCTION

Inductance causes overshoots and undershoots in the signal waveforms, which can adversely affect the signal integrity. For global wires, inductance effects are more severe due to the lower resistance of these lines, which makes the reactive component of the wire impedance comparable to the resistive component, and also due to the presence of significant mutual inductive coupling between wires resulting from longer current return paths. It is shown that the conductors of a circuit system should be regarded as transmission lines for theoretical analysis and practical design in the recent high-speed integrated circuit technology [1].

The design techniques in sub-micron technologies increase the effects of coupling in interconnections [2]. In deep sub-micron technology, the order of capacitive coupling between lines reach to some severe values which signifies that one can't be indifferent to the amplitudes of the noise due to this coupling. Integrated circuit feature sizes continue to scale well below 0.18 microns, active device counts are reaching hundreds of millions [3]. Several factors bound to

the technology contribute to the increase of crosstalk problems: the increase of the number of metal layers [4], the increase in the line thickness, the density of integration and the reduction of the spacing between lines. This set of new challenges is referred to as signal integrity in general. Among all these problem, capacitive coupling induced cross talk is the issue that has been addressed in many literatures. Crosstalk will occur on the chip, on the PCB board, on the connectors, on the chip package, and on the connector cables.

Furthermore, the technology with multi-conductor systems, excessive line-to-line coupling, or crosstalk, can cause two detrimental effects. First, crosstalk will change the performance of the transmission lines in a bus by modifying the effective characteristic impedance and patterns, line-to-line spacing, and switching rates. In this paper, we have proposed a closed form expression for the coupling noise by analyzing the interconnect using RLCG model for ramp input. The major drawback of the proposal made in [5] is that it does not consider the shunt lossy component for estimation of the coupling noise. The proposed model presented in this paper is a generic one in the sense that the model proposed in [5] can be easily derived by just neglecting the shunt lossy component term (i.e. $G=0$).

The rest of the paper is organized as follows: Section 2 discusses the basic theory, transmission line model, crosstalk, glitch and different modes of propagation. Section 3 describes the difference model and the proposed method for noise and delay calculation. Section 4 shows the simulation results. Finally section 5 concludes the paper.

II. BASIC THEORY

A. TRANSMISSION LINE MODEL

Defining the point at which an interconnect may be treated as a transmission line and hence, reflection analysis applied, has no consensus of opinion. A rule of thumb is that when the delay from one end to the other is greater than $\text{risetime}/2$, the line is considered as electrically long. If the delay is less than $\text{risetime}/2$, the line is electrically short. A transmission line [6] can be described at the circuit level using series inductance and resistance combined with shunt capacitance and conductance. An infinitesimal unit length of the

transmission line looks like the circuit as shown in Figure 1. The parameters are defined as follows: R = series resistance per unit length, L = series inductance per unit length, G = shunt conductance per unit length, C = shunt capacitance unit length.

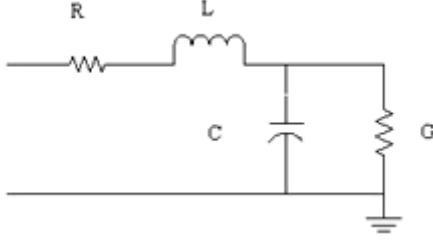


Figure 1. RLCG segment of a transmission line

It is critical to model the transmission path when designing a high-performance, high-speed serial interconnect system. The transmission path may include long transmission lines, connectors, vias and crosstalk from adjacent interconnect. Values for R , L , C , and G are extracted from a given layout, designed in $0.18\mu\text{m}$ technology.

B. CROSS TALK

Crosstalk is the undesired energy imparted to a transmission line due to signals in adjacent lines. The magnitude of the crosstalk induced is a function of rise time, signal line geometry and net configuration (type of terminations, etc.). To overcome the problems faced at high frequency of operation, shielding techniques have been employed [7]. A common method of shielding is to place ground or power lines at the sides of a victim signal line to reduce noise and delay uncertainty.

The crosstalk between two coupled interconnects is often neglected when a shield is inserted, significantly underestimating the coupling noise. The crosstalk noise between two shielded interconnects can produce a peak noise of 15% of V_{DD} in a $0.18\mu\text{m}$ CMOS technology [8]. An accurate estimate of the peak noise for shielded interconnects is therefore crucial for high performance VLSI design. In the complicated multilayered interconnect system, signal coupling and delay strongly affect circuit performances. Thus, accurate interconnect characterization and modelling are essential for today's VLSI circuit design. Two major impacts of cross talk are:

- (i) Crosstalk induces delays, which change the signal propagation time, and thus may lead to setup or hold time failures.
- (ii) Crosstalk induces glitches, which may cause voltage spikes on wire, resulting in false logic behaviour. Crosstalk affects mutual inductance as well as inter-wire capacitance. When the connectors in high speed digital designs are considered, the mutual inductance plays a predominant role compared to the inter-wire capacitance. The effect of mutual inductance is significant in deep submicron technology (DSM) technology since the spacing between two adjacent bus lines is very small. The mutual inductance induces a current from an aggressor line onto a victim line which causes crosstalk between connector lines.

In multi-conductor systems, crosstalk can cause two detrimental effects: first, crosstalk will change the performance of the transmission lines in a bus by modifying the effective characteristic impedance and propagation velocity. Secondly, crosstalk will induce noise onto other lines, which may further degrade the signal integrity and reduce noise margins.

C. GLITCH

Crosstalk Glitch (CTG) is a signal provoked by coupling effects among interconnects lines which have unbalanced drivers and loads [9]. The magnitude of the glitch depends on the ratio of coupling capacitance to that of line to ground capacitance. When a transition signal is applied at a line which has a strong line-driver while stable signals are applied at other lines which have weaker drivers, the stable signals may experience a coupling noise due to the transition of the stronger signal. A glitch may be induced in connector 'j' in which the signal is static, due to neighbouring connector lines in which the signal is varying [10]. This is given by (1).

$$V_{glitch}^j = \sum_j \pm L_{jk} \frac{dj_k}{dt} \forall j \neq k \quad (1)$$

Where L_{jk} represents mutual inductance between j^{th} and k^{th} connector. The sign of the coupled voltage is positive or negative depending upon whether the k^{th} neighbouring connector undergoes a rising or a falling transition.

D. ODD MODE

When two coupled transmission lines are driven with voltages of equal magnitude and 180 degree out of phase with each other, odd mode propagation occurs. The effective capacitance of the transmission line will increase by twice the mutual capacitance, and the equivalent inductance will decrease by the mutual inductance. In Figure 2, a typical transmission line model is considered, where the mutual inductance between aggressor and victim connector is represented as M_{12} . L_1 and L_2 represent the self inductances of aggressor and victim nodes while C_c , C , denote the coupling capacitance between aggressor and victim, self capacitance, respectively.

Assuming $L_1 = L_2 = L_0$, the currents will be of equal magnitude but flow in opposite direction [10]. Thus, the effective inductance due to odd mode of propagation is given by (2).

$$L_{odd} = L_1 - L_2 \quad (2)$$

The magnetic field pattern of the two conductors in odd-mode is shown in Figure 3.

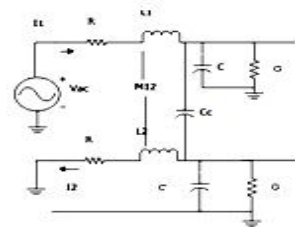


Figure 2. Two Coupled Transmission line model

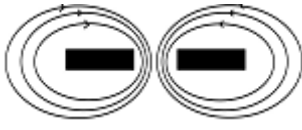


Figure 3. Magnetic Field in Odd Mode

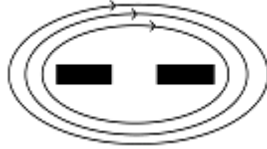


Figure 4. Magnetic Field in Even Mode

E. EVEN MODE

When two coupled transmission lines are driven with voltages of equal magnitude and in phase with each other, even mode of propagation occurs. In this case, the effective capacitance of the transmission line will decrease by the mutual capacitance and the equivalent inductance will increase by the mutual inductance. Thus, in even-mode propagation, the currents will be of equal magnitude and flow in the same direction [10]. The effective inductance, due to even mode of propagation is then given by (3).

III. MODELLING OF CROSS TALK IN RLCG INTERCONNECT

A. DIFFERENCE MODEL

The frequency-domain difference approximation [11] procedure is more general, because it can directly handle lines with arbitrary frequency-dependent parameters or lines characterized by data measured in frequency-domain. The time-domain difference approximation procedure should be employed only if transient characteristics are available. For a single RLCG line, the analytical expressions are obtained for the transient characteristics and limiting values for all the modules of the system and device models. The difference approximation procedure is applied to both the characteristic admittances and propagation functions and the resulting time-domain device models have the same form as the frequency-domain models. The difference approximation procedure involves an approximation of the dynamic part of the system transfer function, with the complex rational series or distorted series. This criterion results in simple and efficient approximation algorithms, and requires a minimal number of the original-function samples to be available, which is important if the line is characterized for delay and crosstalk.

B. ANALYSIS OF CROSSTALK USING DIFFERENCE MODEL

We first consider the interconnect system consisting of single uniform line and ground as shown in Figure 5, and assume the length of the line is d .

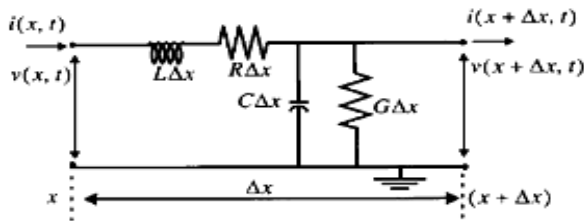


Figure 5. Equivalent Circuit of each Uniform Section

The electrical parameters of each section are $R''x$, $L''x$, $C''x$ and $G''x$ respectively, where R , L , C and G are per-unit length resistance, inductance, capacitance and conductance of the line, respectively.

Using Kirchhoff's Voltage Law (KVL), we can write,

$$v(x, t) = i(x, t)R_{\Delta x} + L_{\Delta x} \frac{di(x, t)}{dt} + v(x + \Delta x, t) \quad (4)$$

Using Kirchhoff's Current Law (KCL), we can write,

$$i(x, t) = G_{\Delta x}v(x + \Delta x, t) + c_{\Delta x} \frac{dv(x + \Delta x, t)}{dt} + i(x + \Delta x, t) \quad (5)$$

Simplifying (4) and (5) and after applying Laplace transform, we get,

$$-\frac{\partial V(x)}{\partial x} = (R + sL)I(x) \quad (6)$$

$$-\frac{\partial I(x)}{\partial x} = (G + sC)V(x) \quad (7)$$

Differentiating (6) and (7) with respect to the x , and after simplifying we get,

$$\frac{\partial^2 V(x)}{\partial x^2} = P^2 V(x) \quad (8)$$

and

$$\frac{\partial^2 I(x)}{\partial x^2} = P^2 I(x) \quad (9)$$

where P is the propagation constant and is defined as,

$$P = \sqrt{(R + sL)(G + sC)} \quad (10)$$

The general solution of (8) is given by

$$V(x) = A_1 e^{-Px} + A_2 e^{Px} \quad (11)$$

where A_1 and A_2 are the constants determined by the boundary conditions. From (8-9) and (11) we get,

$$-\frac{\partial}{\partial x} [A_1 e^{-Px} + A_2 e^{Px}] = (R + sL)I(x) \quad (12)$$

After simplifying we get,

$$I(x) = \frac{1}{Z_0} [A_1 e^{-Px} - A_2 e^{Px}] \quad (13)$$

where Z_0 is the characteristic impedance. Assuming at $x=d$, the termination voltage and current are $V(d) = V_2$ and $I(d) = I_2$, respectively.

$$V_2 = A_1 e^{-Pd} + A_2 e^{Pd} \quad (14)$$

$$I_2 = \frac{1}{Z_0} [A_1 e^{-Pd} - A_2 e^{Pd}] \quad (15)$$

After solving (14) and (15) for A_1 and A_2 we get,

$$A_1 = \frac{1}{2} [V_2 + I_2 Z_0] e^{Pd} \quad (16)$$

$$A_2 = \frac{1}{2} [V_2 - I_2 Z_0] e^{-Pd} \quad (17)$$

Substituting these values of A_1 and A_2 in (11)

$$V(x) = \left[\frac{[V_2 + I_2 Z_0]}{2} e^{P(d-x)} + \frac{[V_2 - I_2 Z_0]}{2} e^{P(x-d)} \right] \quad (18)$$

Now substituting the values of A_1 and A_2 from (16) and (17) in equation (13) we get,

$$I(x) = \frac{1}{Z_0} \left[\frac{[V_2 + I_2 Z_0]}{2} e^{P(d-x)} - \frac{[V_2 - I_2 Z_0]}{2} e^{P(x-d)} \right] \quad (19)$$

Let at $x=0$, $V(x)=V_1$ and $I(x)=I_1$ then from (14) and (18), we can write:

$$V_1 = \cosh(Pd)V_2 + Z_0 \sinh(Pd)I_2 \quad (20)$$

$$I_1 = \frac{1}{Z_0} \sinh(Pd)V_2 + \cosh(Pd)I_2 \quad (21)$$

Since ABCD parameters are defined as

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} \quad (22)$$

Hence, ABCD matrix can be written from (20) and (21) as,

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} \cosh(Pd) & -Z_0 \sinh(Pd) \\ \frac{1}{Z_0} \sinh(Pd) & -\cosh(Pd) \end{bmatrix} \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} \quad (23)$$

The output crosstalk voltage is given by

$$V_2(s) = \frac{V_1(s)}{\cosh(Pd)} \quad (24)$$

For the ramp input voltage we get,

$$V_1(s) = \frac{V_0}{s^2} \quad (25)$$

$$\text{Or, } V_2(s) = \frac{V_0}{s^2 \cosh(\sqrt{(R+sL)(G+sC)})} \quad (26)$$

After simplification, (26) reduces to (27).

$$V_2(s) = \frac{2V_0}{s^2 \left(s + \frac{R}{L} \right) \left(s + \frac{G}{C} \right)} \quad (27)$$

After taking inverse Laplace transform of equation (27), we get,

$$V_0(t) = \frac{-2V_0 LG(LG^2 - R^2 C^2)}{(LG-RC)R^2 G^2} + \frac{2V_0 L C t}{RG} + \frac{2V_0 L^3 C}{R^2 (LG-RC)} e^{-\frac{R}{L}t} - \frac{2V_0 L C^3}{G^2 (LG-RC)} e^{-\frac{G}{C}t} \quad (28)$$

This is the proposed model for noise voltage induced by the aggressor line onto the victim line.

Now we will consider two typical cases of frequency of operation

CASE -1 (For Very Low Frequency)

For very low frequency, where $R \gg \omega L$, (26) reduces to (29).

$$V_2(s) = \frac{V_0}{s^2 \cosh \sqrt{RG}} \quad (29)$$

After taking inverse Laplace transform of (29), we get,

$$v_2(t) = \frac{V_0}{\cosh \sqrt{RG}} tu(t) \quad (30)$$

CASE -2 (For very High Frequency)

For high frequency, where $R \ll \omega L$, (19) reduces to

$$V_2(s) = \frac{V_0}{s^2 \cosh(s\sqrt{LC})} \quad (31)$$

For calculation of the delay expression we consider the 50% rise time when $v_0(t)=0.5v_0$. From (28) we can get the general delay expression.

$$\frac{V_0}{2} = \frac{-2V_0 LG(LG^2 - R^2 C^2)}{(LG-RC)R^2 G^2} + \frac{2V_0 L C t}{RG} + \frac{2V_0 L^3 C}{R^2 (LG-RC)} e^{-\frac{R}{L}t} - \frac{2V_0 L C^3}{G^2 (LG-RC)} e^{-\frac{G}{C}t} \quad (34)$$

After simplification the 50% delay can be derived as,

$$t_{delay} = \sqrt{\frac{LC}{RG}} \quad (35)$$

Equations (28) and (35) represent the proposed model for the crosstalk and the delay.

IV. EXPERIMENTAL RESULTS

The focus of this paper is on the crosstalk and delay estimation of the circuits consisting of lumped elements. Most of the earlier research and reduction techniques consider only capacitive coupling [2, 12]. But in the case of very high frequencies as in GHz scale, inductive crosstalk comes into the important role and it should be included for complete coupling noise analysis. The configuration of circuit for simulation is shown in Figure 2. The high-speed interconnect system consist of two coupled interconnect lines and ground and the length of the lines is $d=10$ mm. The sample dimensions of the cross sections of a minimum sized wire in a $0.18\mu\text{m}$ technology are given in Figure 6.

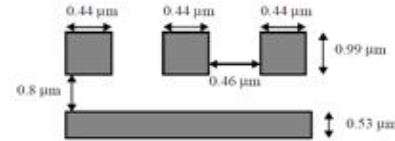


Figure 6. Sample dimensions of cross-sections of minimum sized wire in a $0.18\mu\text{m}$ technology

The extracted values for the parameters R , L , C , and G are given in Table 1. The conductance is a function of frequency, f . The left end of the first line of Figure 2 is excited by 1-V trapezoidal form voltage with rise/fall times 0.5 ns and a pulse width of 1 ns. Other parameters of lumped elements are $R_1=R_2=50$ ohms and $C_1=C_2=1$ pF. R_s , L_s , and C_L are source resistor, source inductor, load capacitor. The operating frequency is taken as 2GHz. Figure-7 shows that

the presence of coupling results in the crosstalk positive glitch (of magnitude 127mV) generated on the victim line and the result is quite comparable to the PSPICE simulation.

TABLE I. RLCG PARAMETERS FOR A MINIMUM- SIZED WIRES IN A $0.18\mu\text{m}$ TECHNOLOGY.

Parameter(s)	Value/m
Resistance(R)	120 kΩ/m
Inductance(L)	270 nH/m
Conductance(G)	15/pS.m
Capacitance(C)	240 pF/m

TABLE II. COMPARISON OF PROPOSED DELAY WITH SPICE

R_s (Ω)	L_s (pH)	C_L (pF)	Spice Delay (ps)	New Ramp Model Delay t_{ramp} (ps)	Step Model Delay t_{step} (ps)
10	0.0246	0.0176	1.56	1.49	1.59
15	0.0246	0.0176	1.64	1.69	1.74
20	0.0246	0.0176	1.74	1.76	1.81
25	0.0246	0.0176	1.83	1.88	1.93

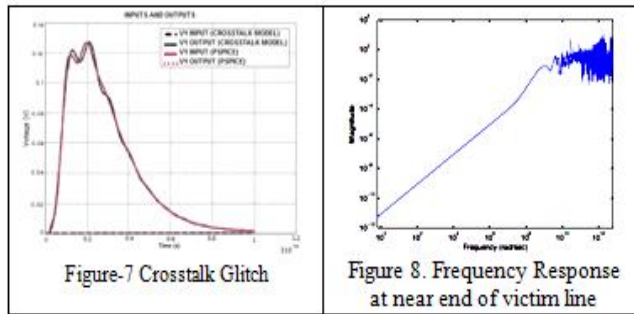


Figure 8. Frequency Response at near end of victim line

From the Figures 8 and 9, we can see that, with the increase of frequency, crosstalk noise becomes larger and larger, so does the attenuation of useful signal. At the range of high frequency, oscillation occurs sharply, which is a bit different from the transfer function of lumped parameter systems. Figure 10 correspond to the waveforms of voltage responses at the both ends of victim line.

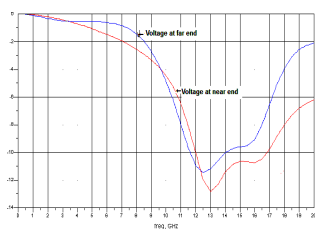
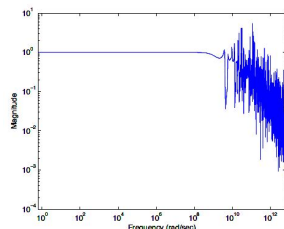


Table-II shows the comparison of the delay obtained from proposed model to that of from SPICE and the step input model [13]. From the result we can see that the error is less than 1%.

CONCLUSIONS

In this paper, explicit models have been proposed for crosstalk noise and delay in high-speed coupled interconnect systems. Result shows that, at low frequencies, the model exhibits a RC behaviour but at high frequencies has a substantially different behaviour due to the effects of inductance. On the basis of Laplace transformation of distributed parameter model deduced in time domain, transfer functions of crosstalk noises are built, and crosstalk noise response is analyzed theoretically. Simulation results demonstrate the validity and correctness of the proposed method.

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